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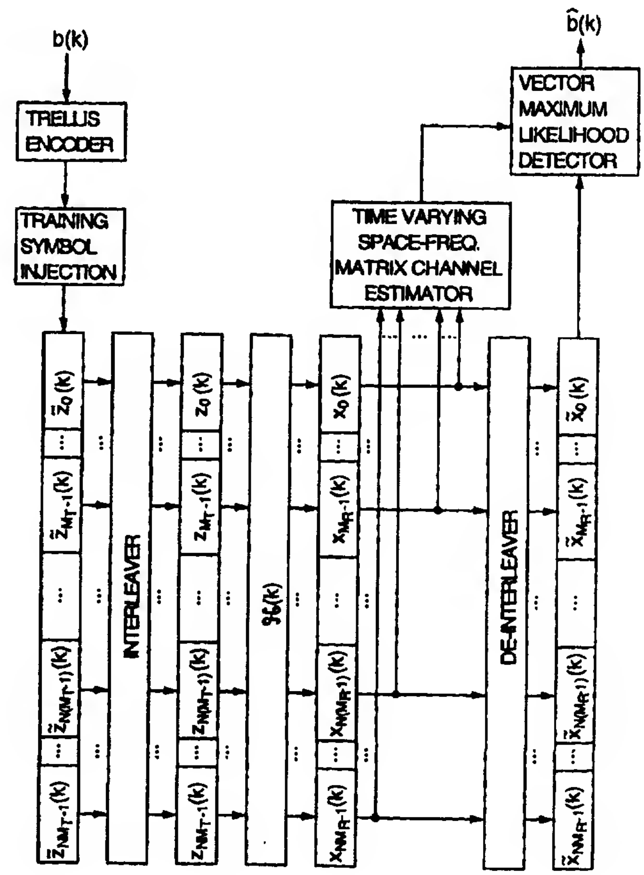
(57) Abstract

In a system and method of digital wireless communication between a base station (B) and a subscriber unit (S), a spatial channel characterized by a channel matrix (H) couples an adaptive array of (MT) antenna elements at the base station (B) with an adaptive array of antenna elements (MR) at the subscriber station (S). The method comprises the use of spatio-temporal coding (TRELLIS ENCODER), training symbols (TRAINING SYMBOL INJECTION), and frequency domain deinterleaving (INTERLEAVER). At the receiver, a matched de-interleaver (DE-INTERLEAVER) transforms the space-frequency sequence back into a serial signal stream. A maximum likelihood detector (VECTOR MAXIMUM LIKELIHOOD DETECTOR) generates the recovered information stream.



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High Capacity Wireless Communication Using Spatio-Temporal Coding

RELATED APPLICATIONS

This application claims priority from U.S. provisional applications 60/025,227 and 60/025,228, both filed 08/29/96. Both applications are hereby incorporated by reference.

FIELD OF THE INVENTION

This invention relates generally to digital wireless communication systems. More particularly, it relates to using antenna arrays by both a base station and a subscriber to significantly increase the capacity of wireless communication systems.

BACKGROUND OF THE INVENTION

Due to the increasing demand for wireless communication, it has become necessary to develop techniques for more efficiently using the allocated frequency bands, i.e. increasing the capacity to communicate information within a limited available bandwidth. This increased capacity can be used to enhance system performance by increasing the number of information channels, by increasing the channel information rates and/or by increasing the channel reliability.

FIG. 1 shows a conventional low capacity wireless communication system. Information is transmitted from a base station B to subscribers $S_1, \ldots, S_9$ by broadcasting omnidirectional signals on one of several predetermined frequency channels. Similarly, the subscribers transmit information back to the base station by broadcasting similar

signals on one of the frequency channels. In this system, multiple users independently access the system through the division of the frequency band into distinct subband frequency channels. This technique is known as frequency division multiple access (FDMA).

A standard technique used by commercial wireless phone systems to increasing capacity is to divide the service region into spatial cells, as shown in FIG. 2. Instead of using just one base station to serve all users in the region, a collection of base stations $B_1, \ldots, B_7$ are used to independently service separate spatial cells. In such a cellular system, multiple users can reuse the same frequency channel without interfering with each other, provided they access the system from different spatial cells. The cellular concept, therefore, is a simple type of spatial division multiple access (SDMA).

In the case of digital communication, additional techniques can be used to increase capacity. A few well known examples are time division multiple access (TDMA) and code division multiple access (CDMA). TDMA allows several users to share a single frequency channel by assigning their data to distinct time slots. CDMA is normally a spread-spectrum technique that does not limit individual signals to narrow frequency channels but spreads them throughout the frequency spectrum of the entire band. Signals sharing the band are distinguished by assigning them different orthogonal digital code sequences. These techniques use digital coding to make more efficient use of the available spectrum.

Wireless systems may also use combinations of the above techniques to increase capacity, e.g. FDMA/CDMA and TDMA/CDMA. Although these and other known techniques increase the capacity of wireless communication systems, there is still a need to further increase system performance. Recently, considerable attention has focused on ways to increasing capacity by further exploiting the spatial domain.

One well-known SDMA technique is to provide the base station with a set of independently controlled directional antennas, thereby dividing the cell into separate sectors, each controlled by a separate antenna. As a result, the frequency reuse in the system can be increased and/or cochannel interference can be reduced. Instead of independently controlled directional antennas, this technique can also be implemented with a coherently controlled antenna array, as shown in FIG. 3. Using a signal processor to control the relative phases of the signals applied to the antenna elements, predetermined beams can be formed in the directions of the separate sectors. Similar signal processing can be used to selectively receive signals only from within the distinct sectors.

In an environment containing a significant number of reflectors (such as buildings), a signal will often follow multiple paths. Because multipath reflections alter the signal directions, the cell space experiences angular mixing and can not be sharply divided into distinct sectors. Multipath can therefore cause cochannel interference between sectors, reducing the benefit of sectoring the cell. In addition, because the separate parts of such a multipath signal can arrive with different phases that destructively interfere, multipath can result in unpredictable signal fading.

In order to avoid the above problems with multipath, more sophisticated SDMA techniques have been proposed. For example, U.S. Pat. No. 5,471,647 and U.S. Pat. No. 5,634,199, both to Gerlach et al., and U.S. Pat. No. 5,592,490 to Barratt et al. disclose wireless communication systems that increase performance by exploiting the spatial domain. In the downlink, the base station determines the spatial channel of each subscriber and uses this channel information to adaptively control its antenna array to form customized beams,

as shown in FIG. 4A. These beams transmit an information signal x over multiple paths so that the signal x arrives to the subscriber with maximum strength. The beams can also be selected to direct nulls to other subscribers so that cochannel interference is reduced. In the uplink, as shown in FIG. 4B, the base station uses the channel information to spatially filter the received signals so that the transmitted signal x' is received with maximum sensitivity and distinguished from the signals transmitted by other subscribers. In this approach the same information signal follows several paths, providing increased spatial redundancy.

In the uplink, there are well known signal processing techniques for estimating the spatial channel from the signals received at the base station antenna array, e.g. by using *a priori* spatial or temporal structures present in the signal, or by blind adaptive estimation. If the uplink and downlink frequencies are the same, then the spatial channel for the downlink is directly related to the spatial channel for the uplink, and the base can use the known uplink channel information to perform transmit beamforming in the downlink. Because the spatial channel is frequency dependent and the uplink and downlink frequencies are often different, the base does not always have sufficient information to derive the downlink spatial channel information. One technique for obtaining downlink channel information is for the subscriber to periodically transmit test signals to the base on the downlink frequency rather than the uplink frequency. Another technique is for the base to transmit test signals and for the subscriber to feedback channel information to the base. If the spatial channel is quickly changing due to the relative movement of the base, the subscriber and/or reflectors in the environment, then the spatial channel must be updated frequently, placing a heavy demand on the system. One method to reduce the required feedback rates is to track only the subspace spanned by the time-averaged channel vector, rather than the instantaneous channel vector. Even with this

reduction, however, the required feedback rates are still a large fraction of the signal information rate.

Although these adaptive beamforming techniques require substantial signal processing and/or large feedback rates to determine the spatial channel in real time, these techniques have the advantage that they can navigate the complex spatial environment and avoid, to some extent, the problems introduced by multipath reflections. As a result, an increase in performance is enjoyed by adaptive antenna array systems, due to their use of the spatial dimension. Note, however, that while the base station antenna array can make efficient use of the spatial dimension by selectively directing the downlink signal to the subscriber S, the uplink signal in these systems is spatially inefficient. Typically, the subscriber is equipped with only a single antenna that radiates signal energy in all directions, potentially causing cochannel interference. These communication systems, therefore, do not make optimal use of the spatial dimension to increase capacity.

OBJECTS AND ADVANTAGES OF THE INVENTION

Accordingly, it is a primary object of the present invention to provide a communication system that significantly increases the capacity and performance of wireless communication systems by taking maximum advantage of the spatial domain. Another object of the invention is to provide computationally efficient coding techniques that make optimal use of the spatial dimensions of the channel. In particular, it is an object of the present invention to provide coding techniques specially adapted for the case of rapidly fading channels where channel state information (CSI) at the transmitter is unknown. These and other objects and advantages will become apparent from the following description and associated drawings.

SUMMARY OF THE INVENTION

These objects and advantages are attained by a method of digital wireless communication that takes maximal advantage of spatial channel dimensions between a base station and a subscriber unit to increase system capacity and performance. Surprisingly, the techniques of the present invention provide an increased information capacity in multipath environments. In contrast, known techniques suffer in the presence of multipath and do not exploit multipath to directly increase system capacity. In brief, the present invention teaches a method of wireless communication using antenna arrays at both the base and subscriber units to transmit distinct information signals over different spatial channels in parallel, thereby multiplying the capacity between the base and the subscriber. In particular, the present invention teaches specific spatio-temporal coding techniques that make optimal use of these additional spatial subchannels in the case of unknown transmitter channel state information.

Generally, the present invention provides a method of digital wireless communication between a base station and a subscriber unit in the case where channel state information is not known by the transmitter. For this purpose a spatio-temporal coding structure that exploits the spatial subchannel capacity is used. In particular, a matrix orthogonal frequency division multiplexing (MOFDM) scheme and a space-frequency trellis coding system is used at the transmitter, and a space-frequency maximum likelihood detector with a channel estimator are used at the receiver. With this relatively simple structure, a MIMO system according to the present invention is able to provide a channel capacity several times greater than can be achieved in a conventional wireless system using OFDM. The inventors also propose an efficient channel estimation algorithm for the time varying MIMO channel.

DESCRIPTION OF THE FIGURES

FIG. 1 shows a low capacity wireless communication system well known in the prior art.

FIG. 2 illustrates a known technique of spatially dividing a service region into cells in order to increase system capacity.

FIG. 3 illustrates the use of beamforming with an antenna array to divide a cell into angular sectors, as is known in the art.

FIGS. 4A and 4B illustrate state-of-the-art techniques using adaptive antenna arrays for downlink and uplink beamforming, respectively.

FIGS. 5A and 5B show the parallel transmission of distinct information signals using spatial subchannels in downlink and uplink, respectively, as taught by the present invention.

FIGS. 6A and 6B are physical and schematic representations, respectively, of a communication channel for a system with multiple transmitting antennas and multiple receiving antennas, according to the present invention.

FIGS. 7A and 7B are block diagrams of the system architecture for communicating information over a multiple-input-multiple-output spatial channel according to the present invention.

DETAILED DESCRIPTION

Although the following detailed description contains many specifics for the purposes of illustration, anyone of ordinary skill in the art will appreciate that many variations and alterations to the following details are within the scope of the invention. Accordingly, the following preferred embodiment of the invention is set forth without any loss of generality to, and without imposing limitations upon, the claimed invention.

As discussed above in relation to FIGS. 4A and 4B, prior art wireless systems employing an adaptive antenna array at the

base station are multiple-input-single-output (MISO) systems, i.e. the channel from the base to the subscriber is characterized by multiple inputs at the transmitting antenna array and a single output at the receiving subscriber antenna. Because these MISO systems can exploit some of the spatial channel, they have an increased capacity as compared to single-input-single-output (SISO) systems that are discussed above in relation to FIGS. 1 and 2. It should be noted that although the MISO systems disclosed in the prior art provide an increase in overall system capacity by spatially isolating separate subscribers from each other, these systems do not provide an increase in the capacity of information transmitted from the base to a single subscriber, or vice versa. As shown in FIGS. 4A and 4B, only one information signal is transmitted between the base and subscriber in both downlink and uplink of a MISO system. Even in the case where the subscriber is provided with an antenna array, the prior art suggests only that this capability would further reduce cochannel interference. Although the overall system capacity could be increased, this would not increase the capacity between the base and a single subscriber.

The present invention, in contrast, is a multiple-input-multiple-output (MIMO) wireless communication system that is distinguished by the fact that it increases the capacity of both uplink and downlink transmissions between a base and a subscriber through a novel use of additional spatial channel dimensions. The present inventors have recognized the possibility of exploiting multiple parallel spatial subchannels between a base station and a subscriber, thereby making use of additional spatial dimensions to increase the capacity of wireless communication. Surprisingly, this technique provides an increased information capacity and performance in multipath environments, a result that is in striking contrast with conventional wisdom.

FIGS. 5A and 5B illustrate a MIMO wireless communication system according to the present invention. As shown in FIG. 5A, a base station B uses adaptive antenna arrays and spatial processing to transmit distinct downlink signals x_1, x_2, x_3 through separate spatial subchannels to a subscriber unit S which uses an adaptive array and spatial processing to receive the separate signals. In a similar manner, the subscriber S uses an adaptive array to transmit distinct uplink signals x'_1, x'_2, x'_3 to the base B over the same spatial subchannels. As the multipath in the environment increases, the channel acquires a richer spatial structure that allows more subchannels to be used for increased capacity.

It is important to note that the simple assignment of the distinct signals to the distinct spatial paths in a one-to-one correspondence, as illustrated above, is only one possible way to exploit the additional capacity provided by the spatial subchannel structure. For example, coding techniques can be used to mix the signal information among the various paths. In addition, the present inventors have developed techniques for coupling these additional spatial dimensions to available temporal and/or frequency dimensions prior to transmission. Although such coupled spatio-temporal coding techniques are more subtle than direct spatial coding alone, they provide better system performance, as will be described in detail below.

It is also important to note that the transmit beamforming at the base requires knowledge of the downlink channel state information. (Similarly, the transmit beamforming at the subscriber requires knowledge of the uplink channel state information. Because the system is symmetric with respect to the base and subscriber, it suffices to discuss one case.) Although downlink channel state information can be fed back to the base from the subscriber, if the channel is rapidly changing, then the demand on the channel capacity to provide real time channel information and the demand on the signal

processing may make it impractical to implement the system under the assumption that transmit channel state information is available. Accordingly, the inventors have developed an MOFDM coding technique to take advantage of the added spatial subchannels even in the case of unknown transmitter channel state information.

In order to facilitate an understanding of the present invention and enable those skilled in the art to practice it, the following description includes a teaching of the general principles of the invention, as well as implementation details. First we develop a compact model for understanding frequency dispersive, spatially selective wireless MIMO channels in the case where the channels are time invariant, and then generalize to the case where the channels vary with time. We then discuss the theoretical information capacity limits of these channels, and propose spatio-temporal coding structures that exploit the spatial subchannel capacity in the case of unknown channel state information. In particular, a matrix orthogonal frequency division multiplexing (MOFDM) scheme is described. In a preferred embodiment a space-frequency trellis coding system is located at the transmitter, and a space-frequency maximum likelihood detector with a channel estimator are located at the receiver. With this relatively simple structure, a MIMO system according to the present invention is able to provide a channel capacity several times greater than can be achieved in a conventional wireless system using OFDM. The inventors also propose an efficient channel estimation algorithm for the time varying MIMO channel.

In its preferred implementations, the present invention makes use of many techniques and devices well known in the art of adaptive antenna arrays systems and associated digital beamforming signal processing. These techniques and devices are described in detail in U.S. Pat. No. 5,471,647 and U.S. Pat. No. 5,634,199, both to Gerlach et al., and U.S. Pat. No.

5,592,490 to Barratt et al., which are all incorporated herein by reference. In addition, a comprehensive treatment of the present state of the art is given by John Livita and Titus Kwok-Yeung Lo in *Digital Beamforming in Wireless Communications* (Artech House Publishers, 1996). Accordingly, the following detailed description focuses upon the specific signal processing techniques which are required to enable those skilled in the art to practice the present invention.

Consider first a time-invariant communication channel for a system with M_T transmitting antennas at a base B and M_R receiving antennas at a subscriber S, as illustrated in FIGS. 6A and 6B. The channel input at a sample time k can be represented by an M_T dimensional column vector

$$\mathbf{z}(k) = [z_1(k), \ldots, z_{M_T}(k)]^T,$$

and the channel output and noise for sample k can be represented, respectively, by M_R dimensional column vectors

$$\mathbf{x}(k) = [x_1(k), \ldots, x_{M_R}(k)]^T,$$

and

$$\mathbf{n}(k) = [n_1(k), \ldots, n_{M_R}(k)]^T.$$

The communication over the channel $\mathbf{H}$ may then be expressed as a vector equation

$$\mathbf{x}(k) = \mathbf{H}\mathbf{z}(k) + \mathbf{n}(k),$$

where the MIMO channel matrix is

$$\mathbf{H} = \begin{pmatrix} h_{1,1} & \ldots & h_{1,M_T} \\ \vdots & & \vdots \\ h_{M_R,1} & \ldots & h_{M_R,M_T} \end{pmatrix}.$$

Each matrix element h_{ij} represents the SISO channel between the i^{th} receiver antenna and the j^{th} transmitter antenna. Due to the multipath structure of the spatial channel, orthogonal

spatial subchannels can be determined by calculating the independent modes (e.g. eigenvectors) of the channel matrix **H**. These spatial subchannels can then be used to transmit independent signals and increase the capacity of the communication link between the base B and the subscriber S.

In the case where the channel matrix **H** is not fixed in time, but changes, it should be represented as a time-dependent matrix, **H**(k). Moreover, because the multipath introduces time delays into the various propagation paths, a spatial decomposition of **H** independent of time will result in temporal mixing of the signals. It is more appropriate, therefore, to perform a more general spatio-temporal analysis of the channel.

Let $\{z_j(n)\}$ be a digital symbol sequence to be transmitted from the j^{th} antenna element, $g(t)$ a pulse shaping function impulse response, and T the symbol period. Then the signal applied to the j^{th} antenna element at time t is given by

$$s_j(t) = \sum_n z_j(n) g(t-nT)$$

The pulse shaping function is typically the convolution of two separate filters, one at the transmitter and one at the receiver. The optimum receiver filter is a matched filter. In practice, the pulse shape is windowed resulting in a finite duration impulse response. We assume synchronous complex baseband sampling with symbol period T. We define n_0 and $(\nu+1)$ to be the maximum lag and length over all paths l for the windowed pulse function sequences $\{g(nT - \tau_l)\}$. To simplify notation, it is assumed that $n_0 = 0$, and the discrete-time notation $g(nT - \tau_l) = g_l(n)$ is adopted.

When a block of N data symbols are transmitted, $N+\nu$ non-zero output samples result. Denoting k as the block index for the k^{th} channel usage, $k(N+\nu)$ is the discrete time index for the

first received sample, and $(k+1)(N+\nu)-1$ is the time index for the last received sample. The composite channel output can now be written as an $M_R\cdot(N+\nu)$ dimensional column vector with all time samples for a given receive antenna appearing in order so that

$$\mathbf{x}(k) = [x_1(k(N+\nu)),\ldots,x_1((k+1)(N+\nu)-1),\ldots,\\ x_{M_R}(k(N+\nu)),\ldots,x_{M_R}((k+1)(N+\nu)-1)]^T,$$

with an identical stacking for the output noise samples $\mathbf{n}(k)$. Similarly, the channel input is an $M_T\cdot N$ dimensional column vector written as

$$\mathbf{z}(k) = [z_1(k(N+\nu)),\ldots,z_1(k(N+\nu)+N-1),\ldots,\\ z_{M_T}(k(N+\nu)),\ldots,z_{M_T}(k(N+\nu)+N-1)]^T,$$

The spatio-temporal communication over the channel $\mathbf{H}(k)$ may then be expressed as a vector equation

$$\mathbf{x}(k) = \mathbf{H}(k)\mathbf{z}(k) + \mathbf{n}(k),$$

where the MIMO time-dependent channel matrix

$$\mathbf{H}(k) = \begin{pmatrix} \mathbf{H}_{1,1}(k) & \ldots & \mathbf{H}_{1,M_T}(k) \\ : & & : \\ \mathbf{H}_{M_R,1}(k) & .. & \mathbf{H}_{M_R,M_T}(k) \end{pmatrix}$$

is composed of SISO sub-blocks $\mathbf{H}_{ij}(k)$.

To clearly illustrate the effect of multipath, the channel can be written as the sum over multipath components

$$H(k) = \sum_{l=1}^{L} \begin{bmatrix} a_{R,1}(\theta_{R,l})\mathbf{I} \\ : \\ a_{R,M_R}(\theta_{R,l})\mathbf{I} \end{bmatrix} \beta_l(k)\mathbf{G}_l \left[a_{T,1}(\theta_{T,l})\mathbf{I} \quad \ldots \quad a_{T,M_T}(\theta_{T,l})\mathbf{I} \right].$$

In this equation, $a_{R,j}(\theta_{R,l})$ is the gain response of the j^{th} receiver array element due to angle of arrival $\theta_{R,l}$ of the l^{th} multipath signal, $a_{T,i}(\theta_{T,l})$ is the gain response of the i^{th} transmitter array element due to angle of departure $\theta_{T,l}$ of the l^{th} multipath signal. $\mathfrak{B}_l(k)$ is the diagonal time varying channel fading parameter matrix given by

$$\mathfrak{B}_l(k) = \mathrm{diag}\ [\ \beta_l(k(N+\nu)),\ldots,\ \beta_l((k+1)(N+\nu)-1)\ ],$$

and the Toeplitz pulse shaping matrix $\mathbf{G}_l$ is given by

$$\mathbf{G}_l = \begin{bmatrix} g_l(0) & 0 & 0 & 0 & \ldots & 0 \\ \vdots & \ddots & & \vdots & & \vdots \\ g_l(\nu) & \ldots & g_l(0) & 0 & & 0 \\ 0 & g_l(\nu) & \ldots & g_l(0) & 0 & 0 \\ \vdots & & \ddots & & \ddots & \\ 0 & & 0 & g_l(\nu) & \ldots & g_l(0) \\ \vdots & & \vdots & & \ddots & \vdots \\ 0 & \ldots & 0 & 0 & 0 & g_l(\nu) \end{bmatrix}.$$

We will now discuss the information capacity for the spatio-temporal channel developed above. The following analysis assumes that the noise $\mathbf{n}(k)$ is additive white Gaussian noise (AWGN) with covariance $\sigma^2\mathbf{I}_{\nu+1}$. Each channel use consists of an N symbol burst transmission and the total average power radiated from all antennas and all time samples is constrained to less than a constant.

Write the singular value decomposition (SVD) of the channel matrix as $\mathbf{H}(k)=\mathbf{V}_H(k)\Lambda_H(k)\mathbf{U}_H^*(k)$, with the n^{th} singular value denoted $\lambda_{H,n}(k)$. Write the spatio-temporal covariance matrix for $\mathbf{z}(k)$ for block index k as $\mathbf{R}_Z(k)$ with eigenvalue decomposition $\mathbf{R}_Z(k)=\mathbf{V}_Z(k)\Lambda_Z(k)\mathbf{U}_Z^*(k)$, and eigenvalues $\lambda_{Z,n}(k)$.

It can be demonstrated that, if the case where the instantaneous channel state information is known at both the transmitter and receiver, then the information capacity for

the time-varying discrete-time spatio-temporal communication channel defined above is given by

$$C = E\left(\sum_{n=1}^{K} \log\left(1 + \frac{\lambda_{Z,n}(k)\,|\lambda_{H,n}(k)|^2}{\sigma^2}\right)\right),$$

where $\lambda_{Z,n}(k)$ is given by the spatio-temporal water-filling solution, $E(\cdot)$ is the expectation operator, and K is the number of finite amplitude singular values in $\mathbf{H}(k)$.

For the case where only the receiver has instantaneous channel state information, it is not possible to adapt the transmitter for each block. Nevertheless, it is possible to find the time invariant transmitter covariance which maximizes the capacity for the worst case channel possibilities. For any given transmitted signal covariance matrix $\mathbf{R}_Z$, the worst case channel would place all of the time average energy in the rank 1 subspace defined by the smallest eigendirection in $\mathbf{R}_Z$. This game theoretic problem leads to a spatially uncorrelated transmitter covariance solution $\mathbf{R}_Z = \frac{P_T}{M_T}\mathbf{I}_{M_T}$, where P_T is the maximum average block transmission power. This transmitter covariance is used for completely unknown point to point channels and broadcast channels. For this case of unknown CSI at the transmitter, it can be shown that a white space-time transmission distribution gives a channel capacity

$$C = E\left(\sum_{n=1}^{K} \log\left(1 + \frac{P_T\,|\lambda_{H,n}(k)|^2}{M_T\sigma^2}\right)\right).$$

By analyzing the ranks of the matrices in the path decomposition of the time varying channel $\mathbf{H}(t)$, it can be demonstrated that the maximum number of finite amplitude parallel spatio-temporal channel dimensions, K, that can be created to communicate over the far field time-varying channel

defined above is equal to min{ $N \cdot L$, $(N+\nu) \cdot M_R$, $N \cdot M_T$ }, where L is the number of multipath components. Thus, multipath is an advantage in far-field MIMO channels. If the multipath is large ($L \gg 1$), the capacity can be multiplied by adding antennas to both sides of the radio link. This capacity improvement occurs with no penalty in average radiated power or frequency bandwidth because the number of parallel channel dimensions is increased. In practice, an adaptive antenna array base station, such as that described by Barratt et al., is modified to implement a coding scheme, as described below, which exploits these additional dimensions. In particular, a signal processor is designed to perform a spatio-temporal transform of information signals in accordance with the above equations so that they may be transmitted through the independent parallel subchannels and decoded by the subscriber.

In constant or slowly varying channels, it is often possible to send training sequences to the receiver and communicate channel state information (CSI) back to the transmitter in a manner that accurately tracks time variations. In such cases, the transmitter can implement a coding solution which approaches the theoretical capacity limits. The MIMO communication problem becomes more difficult when the channel fades rapidly in time as is the case with portable wireless communication in the microwave frequency bands. It then becomes impractical to feed back CSI from the receiver to the transmitter due to the information bandwidth required to update the channel state in real time. It is highly desirable in such cases to have a channel coding technique that exploits the spatial dimension of the MIMO problem without requiring any CSI at the transmitter. Such a coding technique has been devised by the present inventors and is described in detail below.

Given the time-varying channel defined by $\mathbf{H}(t)$, it is theoretically possible to create a coding system consisting of

a spatio-temporal encoder, and a spatio-temporal maximum likelihood decoder. The obvious difficulty with such a system is the complexity of the decoder. The complexity of the spatio-temporal decoder can be greatly reduced, however, by using a matrix orthogonal frequency division multiplexing (MOFDM) structure according to the present invention. The complexity reduction occurs because inter-symbol interference (ISI) is eliminated from each OFDM sub-channel.

The MOFDM channel structure is derived under the assumption that the channel is block time invariant over a block of N+2ν symbol periods. Under this assumption, the channel fading matrix $\mathfrak{B}_l(k)$ can be replaced by the scalar fading variable $\beta_l(k)$. Note that the block time invariant assumption is reasonable provided that the block duration $(N+2\nu)T \ll \Delta_\beta$, where Δ_β is the correlation interval for the channel fading variable. (The correlation interval is defined here as the time period required for the fading parameter time-autocorrelation function to decrese to some fraction of the zero-shift value.)

For MOFDM, N data symbols are transmitted during each channel usage. However, a cyclic prefix is added to the data so that the last ν data symbols form a preample to the N data symbol message block. By discarding the first and last ν data symbols at the receiver and retaining only N time samples at the channel output, the new MIMO channel $\hat{\mathbf{H}}(k)$ has a block cyclic structure:

$$\hat{\mathbf{H}}(k) = \sum_{l=1}^{L} \beta_l(k) \begin{bmatrix} a_{R,1}(\theta_{R,l})\mathbf{I} \\ \vdots \\ a_{R,M_R}(\theta_{R,l})\mathbf{I} \end{bmatrix} \hat{\mathbf{G}}_l \left[a_{T,1}(\theta_{T,l})\mathbf{I} \ \ \ldots \ \ a_{T,M_T}(\theta_{T,l})\mathbf{I} \right].$$

where the cyclic pulse shaping matrix $\hat{\mathbf{G}}(k)$ is given by

$$\hat{\mathbf{G}}_l = \begin{bmatrix} g_l(0) & 0 & \cdots & 0 & g_l(\nu) & \cdots & g_l(1) \\ \vdots & \ddots & \ddots & & \ddots & \ddots & \\ g_l(\nu-1) & \cdots & g_l(0) & 0 & \cdots & 0 & g_l(\nu) \\ g_l(\nu) & g_l(\nu-1) & \cdots & g_l(0) & 0 & \cdots & 0 \\ & & & \ddots & \ddots & & \\ 0 & \cdots & 0 & g_l(\nu) & g_l(\nu-1) & \cdots & g_l(0) \end{bmatrix}.$$

The MOFDM channel model can now be derived as follows. First post multiply $\hat{\mathbf{H}}(k)$ with the $N \cdot M_T \times N \cdot M_T$ block diagonal inverse discrete Fourier transform (IDFT) matrix $\mathbf{F}^{*(M_T)}$ where each diagonal block is the unitary $N \times N$ IDFT matrix $\mathbf{F}^*$. The next step is to premultiply by a similar $N \cdot M_R \times N \cdot M_R$ block diagonal DFT matrix $\mathbf{F}^{(M_R)}$ where the diagonal submatrices $\mathbf{F}$ are $N \times N$ DFT matrices. Pre- and post-multiplication by permutation matrices $\mathbf{P}_R$ and $\mathbf{P}_T$ then gives the decomposition of the channel into discrete discrete Fourier transform (DFT) frequency domain sub-channels $\mathcal{H}_n(k)$, as follows:

$$\mathcal{H}(k) = \sqrt{(N)}\,\mathbf{P}_R \mathbf{F}^{(M_R)} \hat{\mathbf{H}}(k) \mathbf{F}^{*(M_T)} \mathbf{P}_T$$
$$= \begin{pmatrix} \mathcal{H}_1(k) & & 0 \\ & \ddots & \\ 0 & & \mathcal{H}_N(k) \end{pmatrix}$$

Each channel $\mathcal{H}_n(k)$ is independent of the other frequency domain sub-channels. Just as in the case of scalar OFDM, the cyclic prefix allows the large time domain channel to be decomposed into many smaller parallel frequency domain channels. The received vector signal $\mathcal{X}_n(k)$ for each frequency domain spatial sub-channel can then be expressed as

$$\mathcal{X}_n(k) = \mathcal{H}_n(k)\mathcal{Z}_n(k) + \mathcal{N}_n(k),$$

where $\mathcal{Z}_n(k)$ is the subchannel transmitted signal and $\mathcal{N}_n(k)$ is the subchannel noise. A system architecture implementing this channel structure is shown in FIG. 7A.

The spatial sub-channels can also be expressed as

$$\mathcal{H}_n(k) = \sum_{l=1}^{L} \beta_l(k)\ \mathcal{G}_{l,n}\ \mathbf{a}_{R,l}\ \mathbf{a}_{T,l}^{T}$$

where $\mathcal{G}_{l,n}$ is the DFT of the sequence $\{g_l(k)\}$ evaluated at DFT index n. At each frequency index, the DMMT channel is due to a weighted sum over L rank-1 outer products of the frequency-invariant receive and transmit array response vectors. The weighting is determined by the frequency invariant path fading values and the Fourier transform of the delayed pulse shaping function. This reveals a highly structured nature for the time varying space-frequency channel spectrum.

In the case of rapidly fading channels where CSI is not available at the transmitter, the appropriate transmitter distribution is a spatially and temporally white transmitter sequence. Nevertheless, as seen from the above channel decomposition, the use of cyclic signal structures allows the determination of a channel structure that can still be exploited to improve capacity. Therefore, a practical subchannel coding method which approximates a white distribution is desired. Although many variations are possible, the following description is focused on a particularly simple strategy involving a one dimensional trellis coding structure.

In MOFDM, a space-frequency code is transmitted. Given M_T transmitting antenna elements and the MOFDM subchannel decomposition of $\mathcal{H}(k)$, a codeword sequence $\mathbf{c}^{(j)}$ of constraint length $N_c^{(j)}$ can be viewed as q spatial vector code segments transmitted in each of $q = \frac{|\mathbf{c}^{(j)}|}{M_T}$ frequency bins where $|\mathbf{c}^{(j)}|$ is the length of the code sequence. In this embodiment, an information signal b(k) is converted into a code sequence $\mathbf{c}^{(j)}$ by a one dimensional trellis encoder, as shown in FIG. 8. Code

segments of length M_T form a spatial vector code $\mathbf{c}_n^{(j)}$ for a single MOFDM frequency bin indexed by n. After training symbols are injected, frequency domain interleaving is performed by an interleaver in order to distribute consecutive spatial vector code segments among well separated frequency bins. Interleaving allows the system to exploit the frequency diversity of the channel while the spatial coding is a form of spatial diversity.

Each of the M_T symbols in a given spatial vector code segment for a given frequency bin are transmitted from one of the antennas. At the receiver, a matched frequency de-interleaver transforms the space-frequency sequence back into a serial signal stream. A tilde, ~, above a variable is used to denote the signal sequence before interleaving and after de-interleaving operations. Define

$$\mathbf{c}^{(j)} = [\ c_0^{(j)}, \ldots, c_{qM_T-1}^{(j)}\ ]^T$$

as the trellis encoder symbol sequence codeword of length qM_T indexed by j. Further define

$$\tilde{\mathfrak{X}}^{(q)}(k) = [\ \tilde{\mathfrak{X}}_{lM_R}(k), \ \ldots, \ \tilde{\mathfrak{X}}_{lM_R+qM_T-1}(k)\ ]^T$$

as the received de-interleaved signal sequence due to the transmitted code $\mathbf{c}^{(j)}$ where lM_R is the beginning index for the received sequence of length qM_R spanning q space-frequency subchannels. The output sequence due to codeword $\mathbf{c}^{(j)}$ can now be written as

$$\tilde{\mathfrak{X}}^{(q)}(k) = \sqrt{\frac{P_T}{M_T}}\ \tilde{\mathfrak{H}}^{(q)}(k)\mathbf{c}^{(j)} + \tilde{\mathfrak{N}}^{(q)}(k)$$

where

$$\mathfrak{H}^{(q)}(k) = \begin{pmatrix} \tilde{\mathfrak{H}}_{l}(k) & & 0 \\ & \ddots & \\ 0 & & \tilde{\mathfrak{H}}_{l+q}(k) \end{pmatrix}$$

and

$$\sqrt{\frac{P_T}{M_T}}\, \mathbf{c}^{(j)} = \begin{pmatrix} \tilde{\mathfrak{Z}}_{lM_T}(k) \\ \vdots \\ \tilde{\mathfrak{Z}}_{l+qM_T-1}(k) \end{pmatrix}.$$

The additive noise term $\tilde{\mathfrak{N}}^{(q)}(k)$ is still white after the MOFDM channel operations.

For a given spatio-temporal symbol code set

$$\mathbf{C} = \{\ \mathbf{c}^{(1)}, \ldots,\ \mathbf{c}^{(J)}\}$$

the maximum likelihood detector is given by

$$\hat{\mathbf{c}} = \arg \max_{\mathbf{c}(j)} \ P(\mathbf{c}^{(j)} | \tilde{\mathfrak{X}}^{(q)}(k)).$$

FIG. 7B shows such a detector which is used to generate the recovered information stream, $\hat{\mathfrak{b}}(k)$. Given that the receiver noise present in each space-frequency sub-channel is multivariate AWGN, it is known that the equivalent decoder optimization is

$$\hat{\mathbf{c}} = \arg \min_{\mathbf{c}(j)} \ \left|\left|\sqrt{\frac{P_T}{M_T}}\ \tilde{\mathfrak{H}}^{(q)}(k)\mathbf{c}^{(j)} - \tilde{\mathfrak{X}}^{(q)}(k) \ = \ \right|\right|_2^2.$$

This equation can be solved efficiently using a vector Viterbi detector similar to that used in ISI channels. The main difference here is that while there is correlation in the received spatial code segment in each frequency bin, the information across frequency bins is uncorrelated. This allows the metric computation to be pruned back to the number

of states in the trellis encoder at the beginning of each new spatial code segment hypothesis test. It is undesirable for the encoder to posses parallel transitions because this reduces the diversity order of the code to one. Therefore, all of the encoder input bits are fed to the convolutional encoder with rate r and there are there is only one member in each of the cosets.

The inventors have discovered that, given a random Rayleigh channel process with uncorrelated spatial fading and perfect frequency domain interleaving and any code set **C**, the upper bound on the average bit error rate for a $1 \times m$ SIMO channel is larger than the bound for a $M \times mM$ MIMO channel, even though the later transmits data at M times the rate of the former. This remarkable fact reveals some very interesting behavior for the proposed MIMO channel coding structure. Although the data rate for the $M_R = M = M_T$ MIMO channel goes up linearly with M, the probability of error bound is smaller than that for the SISO channel. While the transmitter power for each spatial symbol must be reduced as the number of antennas and spatial sequences are increased, the length of the spatial code segment error vector $\mathbf{e}_n(j1, j2)$ also increases to offset the transmitter power reduction. In addition, while the frequency diversity due to the number of frequency bins spanned by the code error sequence is reduced as M increases, the denominator exponent increases due to spatial diversity. Thus, as M increases, the effects of frequency diversity are replaced by spatial diversity. Furthermore, it is clear that the MIMO system can benefit from additional spatial diversity by setting $M_R > M_T$. The m-order spatial diversity error performance of a $1 \times m$ SIMO channel can be achieved with an $M \times mM$ MIMO channel which will again achieve M times the data rate of the SIMO channel while maintaining lower error probability.

An alternative code design metric for the spatio-temporal coding structure presented in relation to FIGS. 7A and 7B is

suggested by observing that the correct error sequence metric for code design is clearly the product of Euclidean distances for each of the M_T length spatial error vector segments in the error event sequences. This metric is strikingly similar to periodic product distance metrics that are known from other contexts.

An important aspect of the present invention is channel estimation. In fast fading channels, overhead penalties for conventional multi carrier training techniques can be severe. A large number of sub-channels N is desired so that cyclic prefix overhead is minimized. Large N corresponds to long OFDM symbol duration. Long symbol duration, in turn, requires short intervals between training. In conventional channel training procedures, an entire OFDM symbol is dedicated to training, and several data symbols are inserted between training symbols. Thus, a trade-off exists between cyclic prefix and training overhead with conventional channel estimation techniques.

Furthermore, in burst-mode transmission applications such as wireless ATM, if the average data rate for a virtual circuit is low, then the time between ATM packets can be large. In such cases, it is not feasible to use an entire DMT symbol for training since the channel can change substantially between training symbols. What is needed is a training strategy that allows "instantaneous updates" for the channel estimation algorithm. The present inventors have developed a training approach and channel estimation algorithm which injects training information along with data into each OFDM symbol. The channel estimation algorithm exploits the correlation properties of the time varying wireless channel to estimate the spatial channel for each MOFDM frequency domain sub-channel.

Imperfect channel knowledge can have an impact on error probability, and channel estimation noise in the receiver will

limit the performance of a spatio-temporal coding system. The inventors have discovered that the effect of channel estimation errors can be modeled as an increase in the effective noise variance. This noise variance increase is an interesting function of the time varying channel correlation function, the portable velocity, the average channel SNR and the design of the channel estimation algorithm. Thus, proper design of the channel estimator is critical for low error probability communication.

In all that follows, we again invoke the spatially uncorrelated Rayleigh fading condition. Although the spatial fading is uncorrelated, there is correlation in the OFDM frequency and time domains which we wish to exploit. The correlation in the frequency domain arises from the delay limited nature of the channel impulse response. The correlation in the time domain fading arises from the band limited Doppler shifts experienced by physical objects which move in the vicinity of the portable. We desire a channel estimation algorithm that exploits these correlation properties in an optimal manner.

To estimate the matrix channel that exists at a given OFDM frequency index, note that we can simply estimate the M_T column vectors of dimension $1 \times M_R$. Given that the column vectors are assumed to fade independently, an optimal training strategy is to transmit M_T different training sequences from each transmitter antenna and estimate the resulting column vectors without considering the information received during training from the other transmitter antennas. In addition, the uncorrelated spatial fading assumption allows each of the scalar elements in a given channel column vector to be estimated independently. Thus, with M_T training sequences transmitted independently from each antenna, we can estimate M_R independent frequency domain scalar channel entries. Thus, the focus is on a SISO training strategy for the frequency domain sub-channels that exist between one transmitter antenna

and one receiver antenna. The SISO estimation algorithm is directly generalized to the MIMO case by stacking SISO estimates from each receiver antenna into columns and exploiting the cyclic shift properties of the DFT.

Our SISO channel estimation strategy will be to transmit training symbols in several equally spaced OFDM sub-channels with data embedded between training symbols. For a discrete time channel which is delay limited to v+1 finite impulse response terms, v+1 OFDM training sub-channels are sufficient to construct an estimate of all N sub-channels.

A SISO OFDM channel estimation algorithm is now described. The channel frequency domain training symbol sequence is defined as

$$\mathcal{Z}_T = \mathrm{diag}\left[\mathcal{Z}_0, \mathcal{Z}_{\frac{N}{v+1}}, \ldots, \mathcal{Z}_{\frac{vN}{v+1}}\right].$$

By construction, $\mathcal{Z}_T\mathcal{Z}_T^* = P_T\mathbf{I}_{v+1}$.

The channel estimation procedure is as follows.

1. Given n_1 past measurements, the present measurement, and n_2 future measurements of the frequency-domain training sub-channel outputs, form n_1+n_2+1 measurements of the time varying channel impulse response vector $\tilde{\mathbf{h}}^{(v+1)}(k)$ by dividing the received known training symbols into the outputs and then performing the IDFT operation, i.e.

$$\tilde{\mathbf{h}}^{(v+1)}(k) = \sqrt{\frac{v+1}{N}}\,\mathbf{F}_{v+1}^{*}\mathcal{Z}_T^{-1}\mathfrak{X}_T^{v+1}(k),$$

where $\mathfrak{X}_T^{v+1}(k) = \left[\mathfrak{X}_0(k), \mathfrak{X}_{\frac{N}{v+1}}(k), \ldots, \mathfrak{X}_{\frac{vN}{v+1}}(k)\right]^T$ and $\mathbf{F}_{v+1}$ is the v+1 point DFT matrix.

2. Form the channel impulse response estimate $\hat{\mathbf{h}}^{(v+1)}(k)$ by applying an optimal linear MMSE estimation filter independently to each of the impulse response measurements $\tilde{\mathbf{h}}^{(v+1)}(k)$, i.e.

$$\hat{\mathbf{h}}^{v+1}(k) = \left[\mathbf{w}_h^* \otimes \mathbf{I}_{v+1}\right]\begin{bmatrix}\tilde{\mathbf{h}}^{v+1}(k-n_1)\\ \vdots \\ \tilde{\mathbf{h}}^{v+1}(k-n_2)\end{bmatrix},$$

where $\otimes$ denotes the Kronecker product and $\mathbf{w}_h$ is the scalar Weiner filter for $\hat{\mathbf{h}}^{(v+1)}(k)$.

3. Form the complete OFDM channel estimate by zero-padding the channel impulse response estimate and performing an N-point FFT, i.e.

$$\mathfrak{H}^{(N)} = \mathbf{F}_N\,[\ \hat{\mathbf{h}}^{(v+1)}(k),\ 0,\ \ldots,\ 0\ ]^T.$$

Given the iid fading assumption on the channel impulse response terms, the above channel estimation algorithm is optimal within the class of linear MMSE estimators.

To extend the preceding scalar channel analysis methods to estimate the OFDM matrix subchannels, the following procedure is employed. Rather than transmitting v+1 training symbols spaced by $\frac{N}{v+1}$ sub-channels, we transmit v+1 frequency domain sequences, each of length M_T. This training scheme is illustrated in Table 1 where the notation T(n) represents training symbol n and D(k) represents data symbol k.

Table 1

FFT bin	Content
0	T(0)
:	:
M - 1	T(M-1)
M	D(0)
:	:
N/(ν+1) - 1	D(N/(ν+1) - M - 1)
:	:
νN/(ν+1)	T(νM)
:	:
νN/(ν+1) + M - 1	T(M(ν+1) - 1)
νN/(ν+1) + M	D((ν-1)(N/(ν+1)-M))
:	:
N - 1	D(νN/(ν+1)-νM - 1)

In each of the M_T long training sequences, the first symbol is transmitted from the first antenna, the second symbol from the second antenna, and so on. The first column in the frequency domain sub-channel matrix response is then estimated by performing the scalar channel estimation algorithm on each of the M_R antenna outputs associated with the $\nu+1$ subchannels which appear first in the M_T long training sequences, and stacking the scalar estimates into a vector. The other columns of the matrix channel are estimated in a similar manner, with the exception that the final frequency domain estimates $\hat{\mathfrak{H}}(k)$ obtained from the channel estimation algorithm are cyclic-shifted to account for the frequency sub-channel offset for each transmitter antenna training sequence.

Using this technique, for a complex $M_T \times M_R$ DMT channel, only $M_T(\nu+1)$ DMT sub-channels are required for training. The channel overhead loss due to training and the cyclic prefix is

then $\frac{M_T(v+1)+v}{N+v}$. This training technique only requires FFTs and FIR channel estimation filters to implement.

Among the various applications of the present invention, one of particular utility is a wideband wireless ATM local area network for a campus environment. The transmitted digital symbol rate is 10 MHz. The portable terminals are mobile with a maximum velocity of 70 miles per hour. The RF carrier frequency is 5.2 GHz. This application is extremely challenging for conventional equalizer based communication structures due to the large delay spread and extremely high Doppler frequency (+/- 540 Hz). The Doppler shift also makes conventional CDMA approaches difficult due to the required power control loop bandwidth. For these reasons, the application is an ideal candidate for MOFDM. In one embodiment, such a MOFDM system may have 3 transmitter antennas and either 3 or 6 receiver antennas.

Thus, it will be clear to one skilled in the art that the above embodiment may be altered in many ways without departing from the scope of the invention. Accordingly, the scope of the invention should be determined by the following claims and their legal equivalents.

CLAIMS

What is claimed is:

1. A method of digital wireless communication between a base station and a subscriber unit, the method comprising:

space-frequency encoding a plurality of information signals into a sequence of transmitted signal vectors, wherein the transmitted signal vectors have M_T complex valued components and are selected to send the information signals over the a collection of independent spatial subchannels;

transmitting the sequence of transmitted signal vectors over a spatial channel coupling an array of M_T antenna elements at the base station with an array of M_R antenna elements at the subscriber unit;

receiving a sequence of received signal vectors at the subscriber unit, wherein the received signal vectors have M_R complex valued components; and

performing a space-frequency maximum likelihood detection upon the received signal vectors to recover the information signals.

2. The method of claim 1 wherein the encoding step comprises performing matrix orthogonal frequency division multiplexing of the information signals.

3. The method of claim 1 further comprising the step of injecting training sequences into the information signals.

4. The method of claim 1 further comprising adding cyclic prefixes to the coded signal prior to the transmitting step.

5. The method of claim 1 wherein the encoding step is performed in accordance with a spatio-temporal subchannel decomposition of the channel into independent modes.

6. A digital wireless communication system comprising:
a base station comprising a base station antenna array and a base station signal processor coupled to the base station antenna array;
a subscriber unit comprising a subscriber antenna array coupled through a wireless channel to the base station antenna array and a subscriber signal processor coupled to the subscriber antenna array;
wherein the base station signal processor encodes downlink signal information by matrix orthogonal frequency division multiplexing the signal information; and
wherein the subscriber signal processor decodes the downlink signal information by vector maximum likelihood detection and space-frequency matrix channel estimation.

7. The system of claim 6 wherein the base station signal processor performs interleaving of the signal information and wherein the subscriber signal processor performs de-interleaving.

8. The system of claim 6 wherein the base station signal processor performs trellis encoding of the signal information.

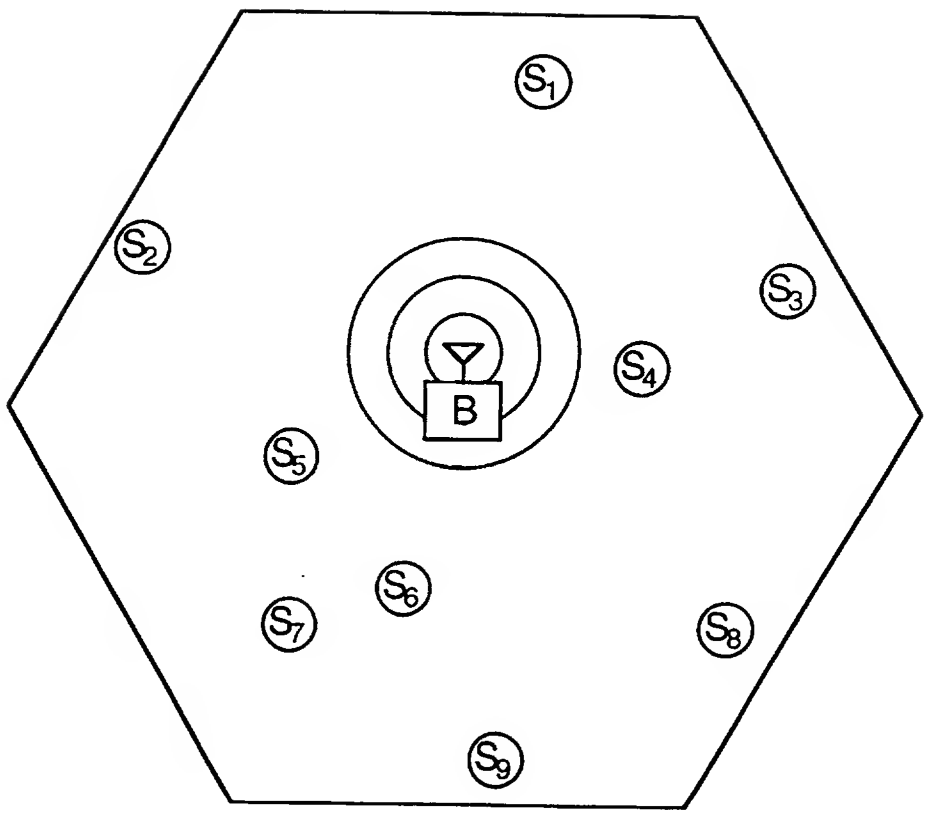


FIG. 1
(PRIOR ART)

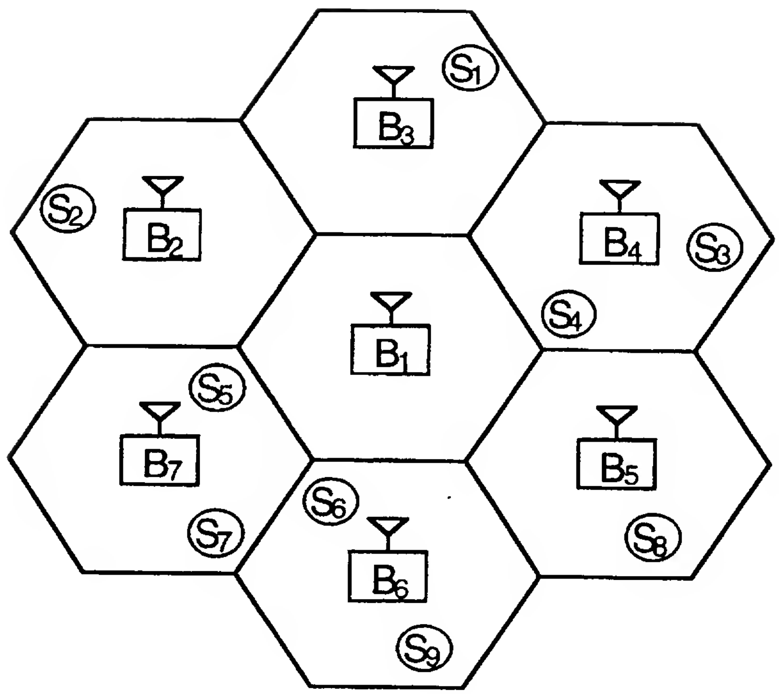


FIG. 2
(PRIOR ART)

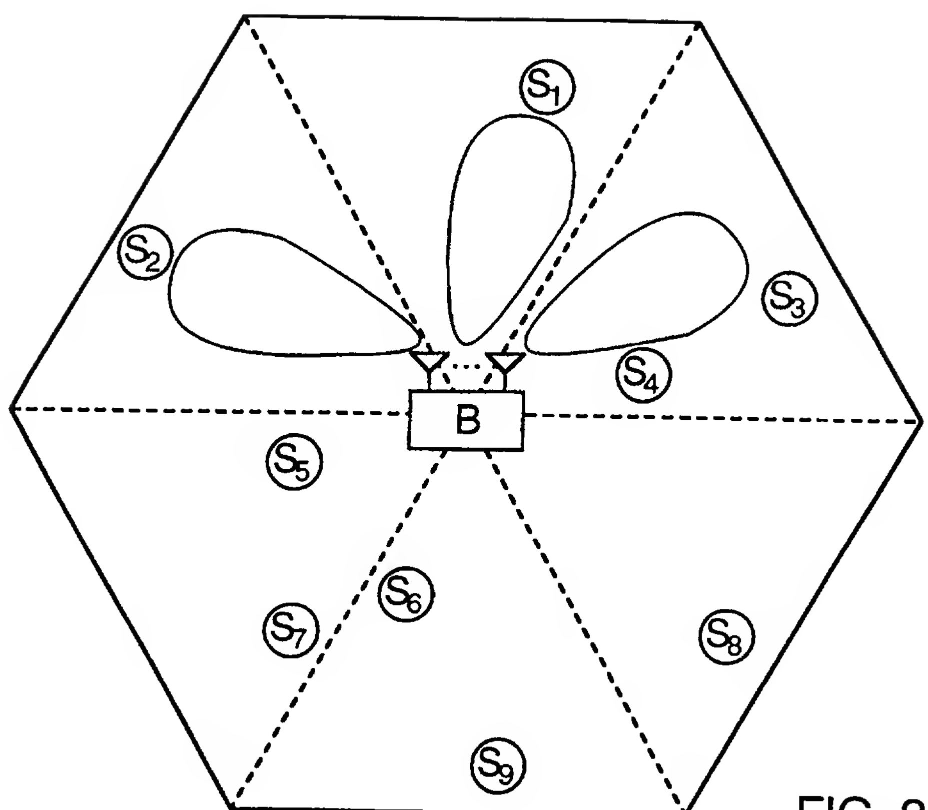


FIG. 3
(PRIOR ART)

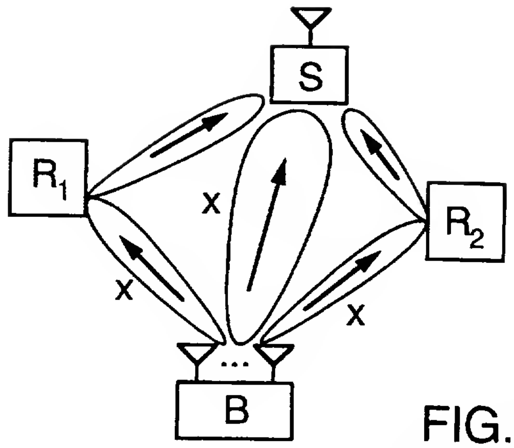


FIG. 4A
(PRIOR ART)

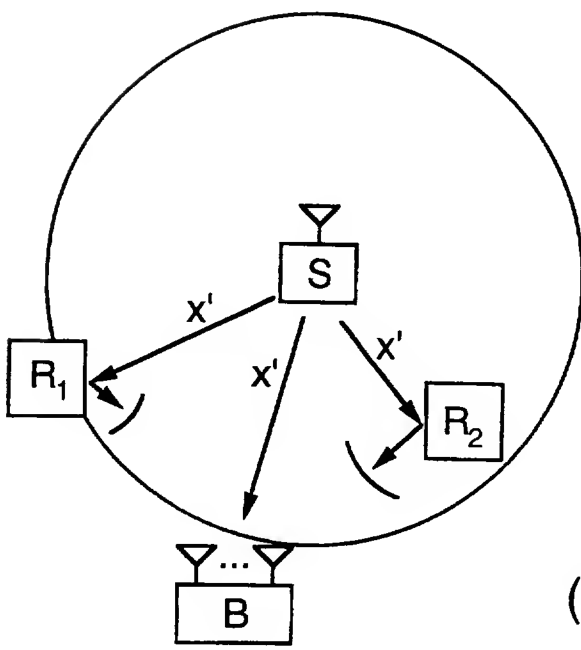


FIG. 4B
(PRIOR ART)

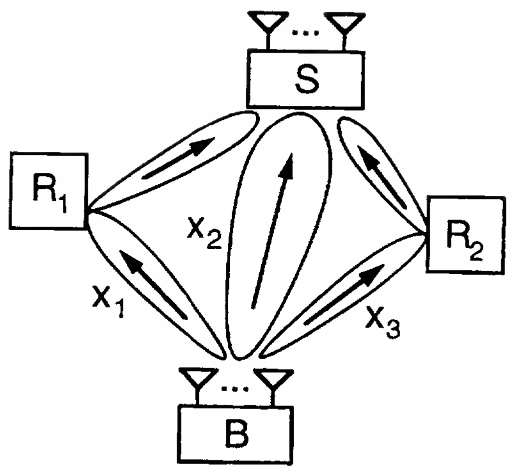


FIG. 5A

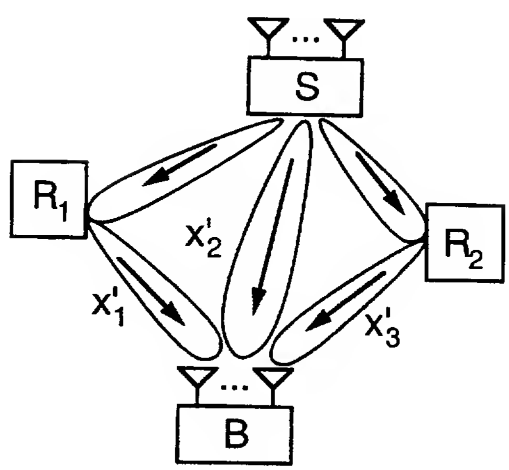


FIG. 5B

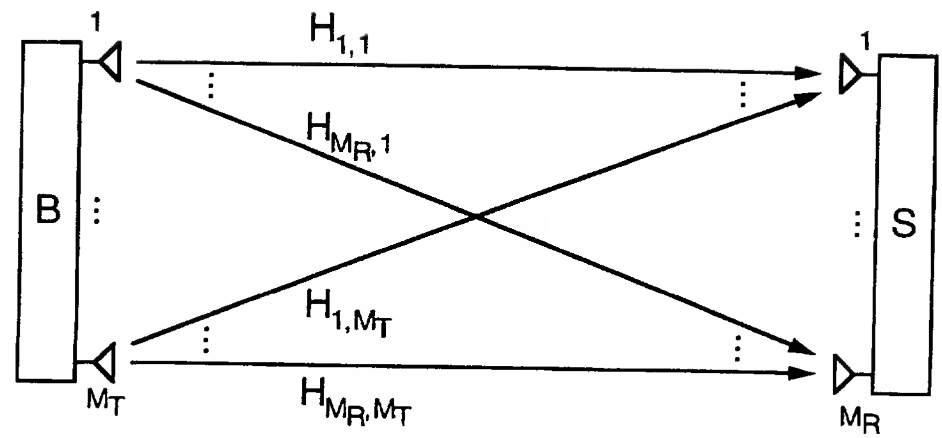

B
S
1
1
M_T
M_R
$H_{1,1}$
$H_{M_R,1}$
H_{1,M_T}
H_{M_R,M_T}

FIG. 6A

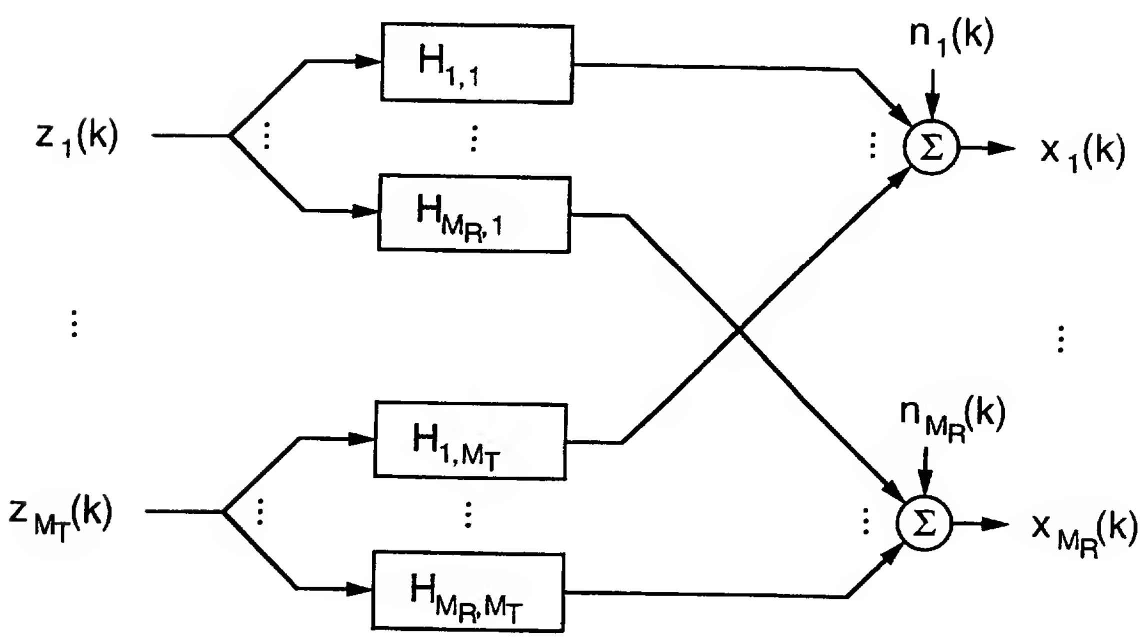
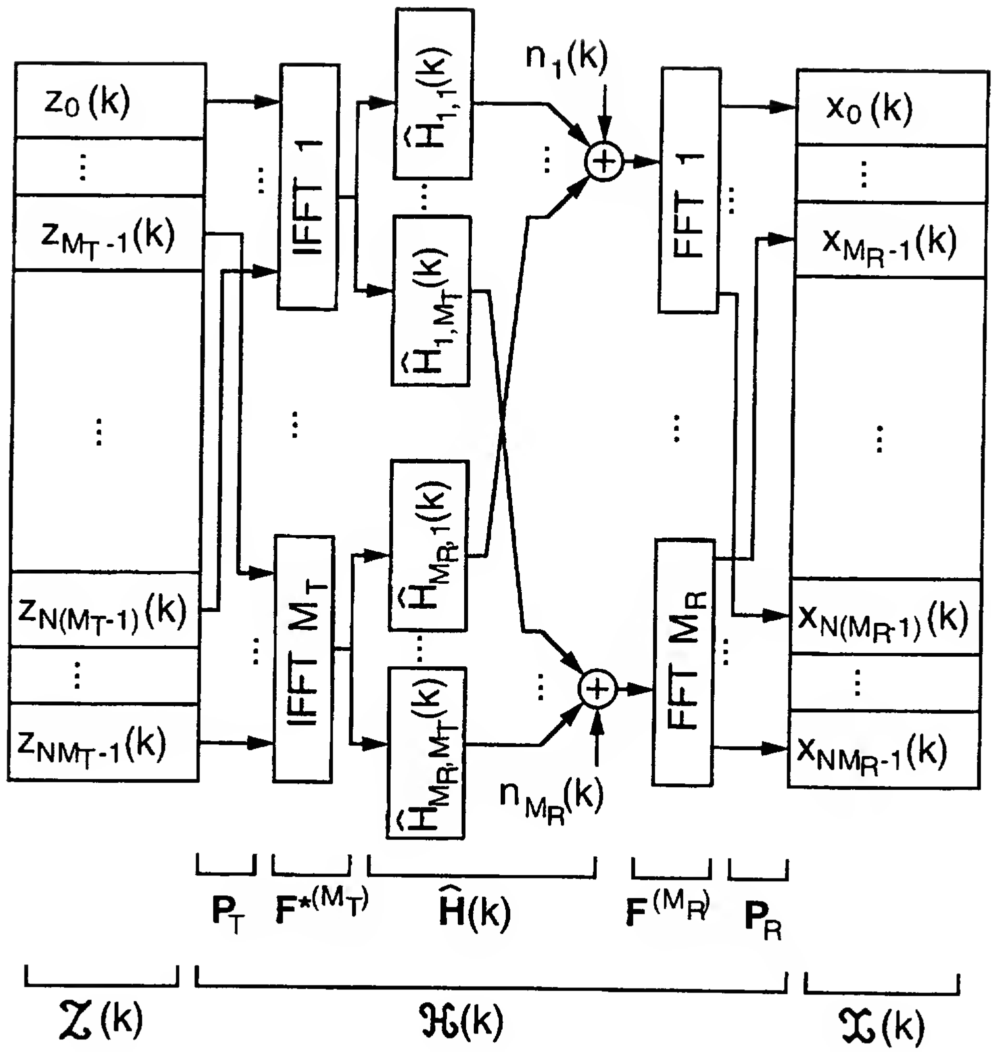

$z_1(k)$
$z_{M_T}(k)$
$H_{1,1}$
$H_{M_R,1}$
H_{1,M_T}
H_{M_R,M_T}
$n_1(k)$
$n_{M_R}(k)$
Σ
Σ
$x_1(k)$
$x_{M_R}(k)$

FIG. 6B



$z_0(k)$
$z_{M_T-1}(k)$
$z_{N(M_T-1)}(k)$
$z_{NM_T-1}(k)$
IFFT 1
IFFT M_T
$\hat{H}_{1,1}(k)$
$\hat{H}_{1,M_T}(k)$
$\hat{H}_{M_R,1}(k)$
$\hat{H}_{M_R,M_T}(k)$
$n_1(k)$
$n_{M_R}(k)$
FFT 1
FFT M_R
$x_0(k)$
$x_{M_R-1}(k)$
$x_{N(M_R-1)}(k)$
$x_{NM_R-1}(k)$
$\mathbf{P}_T$
$\mathbf{F}^{*(M_T)}$
$\hat{\mathbf{H}}(k)$
$\mathbf{F}^{(M_R)}$
$\mathbf{P}_R$
$\mathfrak{Z}(k)$
$\mathfrak{H}(k)$
$\mathfrak{X}(k)$

FIG. 7A

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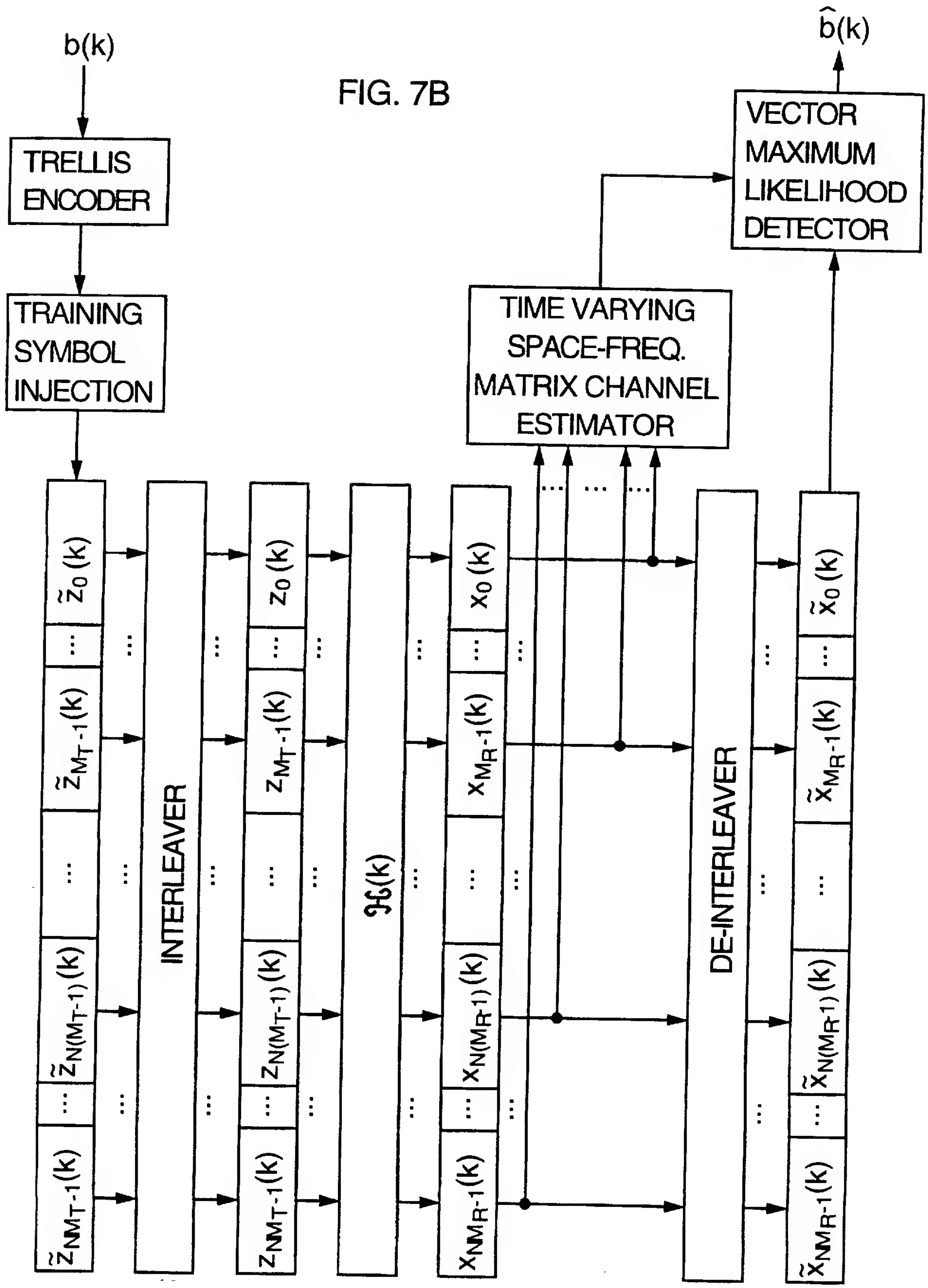
FIG. 7B


b(k)
TRELLIS ENCODER
TRAINING SYMBOL INJECTION
INTERLEAVER
$\mathcal{H}(k)$
TIME VARYING SPACE-FREQ. MATRIX CHANNEL ESTIMATOR
DE-INTERLEAVER
VECTOR MAXIMUM LIKELIHOOD DETECTOR
$\hat{b}(k)$
$\tilde{z}_0(k)$
$\tilde{z}_{M_T-1}(k)$
$\tilde{z}_{N(M_T-1)}(k)$
$\tilde{z}_{NM_T-1}(k)$
$z_0(k)$
$z_{M_T-1}(k)$
$z_{N(M_T-1)}(k)$
$z_{NM_T-1}(k)$
$x_0(k)$
$x_{M_R-1}(k)$
$x_{N(M_R-1)}(k)$
$x_{NM_R-1}(k)$
$\tilde{x}_0(k)$
$\tilde{x}_{M_R-1}(k)$
$\tilde{x}_{N(M_R-1)}(k)$
$\tilde{x}_{NM_R-1}(k)$